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PROVISIONAL APPLICATION FOR PATENT COVER SHEET

This is a request for filing a PROVISIONAL APPLICATION FOR PATENT under 37 CFR 1.53 (c).

INVENTOR(S)

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☐ *Additional inventors are being named on the ___ separately numbered sheets attached hereto*

TITLE OF THE INVENTION (280 characters max)

CONSTRAINED OPTIMIZATION BASED MIMO LMMSE-SIC RECEIVER FOR CDMA DOWNLINK

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ENCLOSED APPLICATION PARTS ***(check all that apply)***

☑ Specification *Number of Pages*	23	☐ Small Entity Statement	
☑ Drawing(s) *Number of Sheets*	8	☑ Other (specify)	Abstract - 1pp Return Postcard_____

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Respectfully submitted,

SIGNATURE [signature]

Date 2/27/04

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Docket Number: NC17729P

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USE ONLY FOR FILING A PROVISIONAL APPLICATION FOR PATENT

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2. EXTRA CLAIM FEES FOR UTILITY AND REISSUE

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1202	18	2202	9	Claims in excess of 20
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1052	50	2052	25	Surcharge - late provisional filing fee or cover sheet	
1053	130	1053	130	Non-English specification	
1812	2,520	1812	2,520	For filing a request for *ex parte* reexamination	
1804	920*	1804	920*	Requesting publication of SIR prior to Examiner action	
1805	1,840*	1805	1,840*	Requesting publication of SIR after Examiner action	
1251	110	2251	55	Extension for reply within first month	
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1253	950	2253	475	Extension for reply within third month	
1254	1,480	2254	740	Extension for reply within fourth month	
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1401	330	2401	165	Notice of Appeal	
1402	330	2402	165	Filing a brief in support of an appeal	
1403	290	2403	145	Request for oral hearing	
1451	1,510	1451	1,510	Petition to institute a public use proceeding	
1452	110	2452	55	Petition to revive - unavoidable	
1453	1,330	2453	665	Petition to revive - unintentional	
1501	1,330	2501	665	Utility issue fee (or reissue)	
1502	480	2502	240	Design issue fee	
1503	640	2503	320	Plant issue fee	
1460	130	1460	130	Petitions to the Commissioner	
1807	50	1807	50	Processing fee under 37 CFR 1.17(q)	
1806	180	1806	180	Submission of Information Disclosure Stmt	
8021	40	8021	40	Recording each patent assignment per property (times number of properties)	
1809	770	2809	385	Filing a submission after final rejection (37 CFR 1.129(a))	
1810	770	2810	385	For each additional invention to be examined (37 CFR 1.129(b))	
1801	770	2801	385	Request for Continued Examination (RCE)	
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SUBMITTED BY				(Complete *(if applicable)*)	
Name *(Print/Type)*	Steven A. Shaw	Registration No. *(Attorney/Agent)*	39,368	Telephone	972-894-6173
Signature	[signature]			Date	2/27/04

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Provisional Application Papers of Jianzhong Zhang, P.R.China;
Balaji Raghothaman; India;
Giridhar Mandyam; US

CONSTRAINED OPTIMIZATION BASED MIMO LMMSE-SIC RECEIVER FOR CDMA DOWNLINK

BACKGROUND

[0001] **References:**

[1] E. Telatar, "Capacity of multi-antenna gaussian channels," *Bell Labs Technical Journal*, June 1995.

[2] P. F. Driessen and G. J. Foschini, "On the capacity formula for multiple input - multiple output wireless channels," *IEEE Transcations on Communications*, vol. 47, pp. 173–176, Feb. 1999.

[3] G. J. Foschini, "Layered space-time architecture for wireless communications in a fading enviroment when using multi-element antennas," *Bell Labs Technical Journal*, pp. 41–59, 1996.

[4] P. W. Wolniansky, G. J. Foschini, G. D. Golden, and R. A. Valenzuela, "V-BLAST: Architecture for realizing very high data rates over the rich-scattering wireless channel," in *Proceedings of ISSSE*, 1998.

[5] G. J. Foschini, G. D. Golden, R. A. Valenzuela, and P. W. Wolnianski, "Simplified processing for high spectral efficiency wireless communication employing multi-element arrays," *IEEE Journal on Selected Areas of Communications*, vol. 17, pp. 1841–1852, November 1999.

[6] M. K. Varanasi and T. Guess, "Optimal decision feedback multiuser equlization with successive decoding

achieves the total capacity of the gaussian multiple-access channel," in *Proceedings of Asilomar Conference*, November 1997.

[7] M. K. Varanasi, "Decision feedback multiuser detection: A systematic approach," *IEEE Transcations on Information Theory*, vol. 45, pp. 219–240, Jan. 1999.

[8] A. Klein, "Data detection algorithms specially designed for the downlink of CDMA mobile radio systems," in *Proc. of VTC 97*, (Phonenix, Arizona), pp. 203–207, 1997.

[9] I. Ghauri and D. T. M. Slock, "Linear receivers for the DS-CDMA downlink exploiting orthogonality of spreading sequences," in *Proc. of 32nd Asilomar Conference*, pp. 650–654, 1998.

[10] S. Werner and J. Lilleberg, "Downlink channel decorrelation in CDMA systems with long codes," in *Proc. of 49th VTC*, pp. 1614–1617, 1999.

[0002] This invention pertains in general to communication systems. More particularly, embodiments of the invention pertain transmit diversity and Multiple-In, Multiple-Out (MIMO) transmission and receiving methods for multiple antenna technology of Code Division Multiple Access (CDMA) type systems.

[0003] Inter-chip interference (ICI) is a result of the multipath frequency selective channel in the CDMA downlink. The presence of ICI destroys the orthogonality of the Walsh spreading codes at mobile terminals. The challenge to the receiver design is even greater for a MIMO system in CDMA downlink. The receiver has to combat both the ICI and the co-channel interference (CCI) to achieve reliable communication. Therefore, interference cancellation at the mobile stations is an effective means of improving the receiver performance and link capacity.

[0004] For a Single-In, Single-Out (SISO) or Single-In, Multiple-Out (SIMO) system, there are two types of linear equalizers: non-adaptive and adaptive equalizers. Non-adaptive equalizer usually requires matrix inversion and is therefore computationally expensive [1-4], especially when the coherence time of the channel is short and the equalizer requires frequent update. Adaptive equalizer is less computationally involved since it does not require direct matrix inversion, but it is not as robust as the non-adaptive equalizers since its performance is sensitive to the choice of parameters such as step size, initialization, etc [5-7]. To overcome these difficulties, in [8] an FFT-based linear equalization algorithm was proposed for SISO/SIMO systems that provides a good tradeoff between complexity and performance. The algorithm approximates the non-adaptive LMMSE algorithm by exploiting the banded Toeplitz structure of the correlation matrix of received signal vector. Another attractive alternative is the Kalman filtering approach proposed in [9] where an interesting state-space model is established to facilitate the application of the recursive Kalman filter. This approach outperforms the LMMSE approach at higher complexity.

[0005] For a MIMO system, it has also been shown in [9] that both conventional LMMSE algorithm and the Kalman filter algorithm can be extended to the MIMO situation. In these MIMO solutions, both the ICI and CCI are suppressed in the linear equalization phase of the receiver. However, these solutions require complicated signal processing whose computational complexities may be beyond practical limits of the current hardware implementation. An FFT-based low-complexity MIMO LMMSE-FFT algorithm was recently proposed in [10] to overcome these difficulties. On the other hand, attempts to combine the non-linear decision feedback interference cancellation with the LMMSE equalization are also found in the literature, for example, [11]. However, these algorithms performs the decision feedback directly at the received signal, and thus requires the impractical assumption that all the active Walsh codes are known at the mobile receiver in order to reconstruct the transmitted chip sequences.

SUMMARY

[0006] A system according to embodiments of this invention provides a multiple transmit antenna, multiple receive antenna (MIMO) receiver design for the communication downlinks such as those used in CDMA technology. Two algorithms referred to as the MIMO LMMSE-FFT and MIMO LMMSE-SIC (Successive Interference Cancellation) algorithms, are described in detail. In embodiments of the invention, the interference cancellation step is achieved without the impractical assumption of the knowledge of all the active Walsh codes in the systems.

[0007] These and other features, aspects, and advantages of embodiments of the present invention will become apparent with reference to the following description in conjunction with the accompanying drawings. It is to be understood, however, that the drawings are designed solely for the purposes of illustration and not as a definition of the limits of the invention.

BRIEF DESCRIPTION OF THE DRAWINGS

Figure 1 is illustrative of a MIMO-CDMA system.

Figure 2 shows transmit signaling at antenna m of the system of figure 1.

Figure 3 is a block diagram showing the MIMO-LMMSE algorithm.

Figure 4 is a block diagram showing a LMMSE-SIC receiver.

Figure 5 is an illustration of a mixed traffic 1X EV-DV system.

Figure 6 is a illustration of the conditional mean estimator.

Figure 7 is a graph showing simulation results for the MIMO LMMSE-SIC algorithm.

Figure 8 is a graph showing simulation results for the MIMO LMMSE-SIC algorithm with 2 iterations.

DETAILED DESCRIPTION

[0008] Embodiments of this invention provide a multiple transmit antenna, multiple receive antenna (MIMO) receiver design for the communication downlinks such as those used in CDMA technology.

[0009] Multiple transmit, multiple receive antenna MIMO systems offer potential for realizing high spectral efficiency of a wireless communications system. The information theoretic studies in [1] and [2] establish that in an independent at-fading channel environment, the capacity of such an MIMO system increases linearly with the number of antennas. In [3], a practical MIMO configuration, a Bell Labs Layered Space-Time (BLAST) system, is deployed to realize this high spectral efficiency for a narrow-band TDMA system. A simpler form of BLAST, the vertical BLAST (V-BLAST) is later proposed in [4, 5] to simplify the transceiver BLAST design. In V-BLAST, independent parallel data sequences are transmitted and a successive co-channel interference CCI cancellation/data detection algorithm is used for efficient multi-symbol detection. Note that the successive detection algorithm is also similar to those proposed in the CDMA multi-user detection literature, for example [6] [7].

[0010] In the downlink of a CDMA like system, most of the research up-to-date focus on the application of the advanced signal processing techniques to offset the performance degradation due to the loss of Walsh code orthogonality caused by the frequency selective multipathchannel. Channel equalization is a promising means of improving the receiver performance in a frequency selective CDMA downlink. Current research encompasses two favors of linear equalization, namely non-adaptive linear equalization [8] [9] [10] [11] and adaptive linear equalization [12] [13]. Non-adaptive linear equalizers usually assumes “piece-wise” stationarity of the channel, and designs the equalizer according to some optimization criteria such as Linear Minimum Mean Square Error (LMMSE) or zero-forcing, which in general leads to solving a system of linear equations by matrix inversion. This can be computationally expensive, especially when the coherence time of the

channel is short and the equalizers have to be updated frequently. On the other hand, adaptive algorithms solve the similar LMMSE or zero-forcing optimization problems by means of stochastic gradient algorithms and avoid direct matrix inversion. Although computationally more manageable, the adaptive algorithms are less robust since their convergence behavior and performance depend on the choices of parameters such as step size. To overcome these difficulties, in [14] an FFT-based linear equalization algorithm was proposed that provides a good trade off between complexity and performance. The algorithm approximates the non-adaptive LMMSE algorithm by exploiting the banded Toeplitz structure of the correlation matrix of received signal vector. Another attractive alternative is the Kalman filtering approach proposed in [15], where an interesting 2 state-space model is established to facilitate the application of the recursive Kalman filter. This approach outperforms the LMMSE approach at a slightly higher complexity.

[0011] Applying MIMO configuration to the CDMA downlink presents significant challenge to the receiver design, as the receiver has to combat both the inter-chip interference (ICI) and the CCI in order to achieve reliable communication. It has been shown in [15] that both conventional LMMSE algorithm and the Kalman filter algorithm can be extended to the MIMO system. The attempts to combine the non-linear decision feedback interference cancellation with the LMMSE equalization are also found in the literature, for example, [16]. However, these algorithms performs the decision feedback directly at the received signal, and thus requires the impractical assumption that all the active Walsh codes are known at the mobile receiver in order to reconstruct the transmitted chip sequences.

[0012] In embodiments of the invention, we propose a MIMO LMMSE-SIC algorithm combines the idea of linear equalization and non-linear decision feedback successive interference cancellation to improve the performance of a conventional linear MIMO LMMSE equalizer. One key advantage of the proposed MIMO LMMSE-SIC algorithm is that the interference cancellation step is achieved without the impractical assumption of the knowledge of all the active Walsh codes in the systems, unlike many other interference cancellation based algorithms found in the literature.

Noteworthy, in the proposed MIMO LMMSE-SIC algorithm, we have also incorporated the so-called conditional mean estimator to provide soft decisions in the decision feedback process. The simulation results suggest that the soft decisions alleviate the error propagation problem in the SIC associated with hard decision feedbacks.

MIMO Signal Model for CDMA Downlink

[0013] Consider an M transmit antenna, N receive antenna MIMO CDMA system as illustrated in Figure 1. Note that since Nokia's MIMO proposal on the transmitter side is yet to be finalized, in the receiver algorithm development process we have assumed a rather simple "serial to parallel split" transmit multiplexing, in order to make our receiver solutions general enough for all possible MIMO transmit multiplexing methods. The modulated symbol streams are split at the transmitter into M sub-streams before transmitted across the M transmit antennas.

[0014] As shown in Figure 2, the signal model at the m_{th} transmit antenna is given as follows, assuming K active Walsh codes in the system:

$$d_m(i) = c(i) \sum_{k=1}^{K} \sum_{m} \alpha_k a_{k,m}(j) s_k(i - jG) \tag{1}$$

where i, j, m and k are chip, symbol, transmit antenna and spreading code indices. The base station scrambling code is denoted by $c(i)$. Meanwhile, α_k stands for the power assigned to spreading code k (same for all antennas), $a_{k,m}$ is the information symbol sequence for spreading code k at antenna m and s_k is the k_{th} spreading code. Note that in this model we have implicitly assumed that the same set of Walsh codes are used across all the transmit antennas.

[0015] The transmitted signals propagates through the MIMO multipath fading channel denoted by $\boldsymbol{H}_0, \ldots, \boldsymbol{H}_L$, where each matrix is of dimension $N\Delta \times M$, where Δ denotes number of samples per chip. The signal model at the receive antennas are thus given by the following equation, after stacking up the received samples across all the receive antennas for the i_{th} chip interval.

$$\mathbf{y}_i = \sum_{l=0}^{L} \mathcal{H}_l \mathbf{d}_{i-l} + \mathbf{n}_i. \tag{2}$$

Note that $\mathbf{y}_i = [\mathbf{y}_{i,1}^T, ..., \mathbf{y}_{i,N}^T]^T$ is of length $N\Delta$, and each small vector $\mathbf{y}_{i,n}$ includes all the temporal samples within the *ith* chip interval. Meanwhile, L is the channel memory length, $\mathbf{d}_{i-l} = [d_1(i-l), ..., d_M(i-l)]^T$ is the transmitted chip vector at time $i - l$, and $\mathbf{n}_i$ is the $(N\Delta) \times 1$ dimensional white Gaussian noise vector with $\mathbf{n}_i \sim \mathcal{N}(\mathbf{0}, \sigma^2\mathbf{I})$. Note that σ^2 denotes noise variance and **I** is the identity matrix. Furthermore, in order to facilitate the discussion on the LMMSE receiver, we stack up a block of $2F + 1$ received vectors:

$$\mathbf{y}_{i+F:i-F} = \mathbf{H}\mathbf{d}_{i+F:i-F-L} + \mathbf{n}_{i+F:i-F} \tag{3}$$

where $2F + 1$ is the length of the LMMSE equalizing filter and

$$\begin{aligned}
\mathbf{y}_{i+F:i-F} &= [\mathbf{y}_{i+F}^T, ..., \mathbf{y}_{i-F}^T]^T, \quad ((2F+1)N\Delta \times 1) \\
\mathbf{n}_{i+F:i-F} &= [\mathbf{n}_{i+F}^T, ..., \mathbf{n}_{i-F}^T]^T, \quad ((2F+1)N\Delta \times 1) \\
\mathbf{d}_{i+F:i-F-L} &= [\mathbf{d}_{i+F}^T, ..., \mathbf{d}_{i-F-L}^T]^T, \quad ((2F+L+1)M \times 1)
\end{aligned}$$

$$\mathbf{H} = \begin{bmatrix} \mathcal{H}_0 & \cdots & \mathcal{H}_L & & \\ & \ddots & & \ddots & \\ & & \mathcal{H}_0 & \cdots & \mathcal{H}_L \end{bmatrix}, \quad ((2F+1)N\Delta \times (2F+L+1)M)$$

where the dimensions of the matrices are given next to them. Note that to keep the notation more intuitive, we keep the subscripts at a "block" level. For instance, $\mathbf{y}_{i+F:i-F}$ is the vector that contains blocks $\mathbf{y}_{i+F}, ..., \mathbf{y}_{i-F}$ where each block is a vector of size $N\Delta \times 1$.

MIMO LMMSE Chip-Level Equalization

[0016] The block diagram of the MIMO receiver with chip-level equalizer is shown in Figure 3. It is essentially a straightforward extension of the LMMSE chip equalization initially designed for a SISO system [8]-[15]. After the chip-level equalizer, the orthogonality of the Walsh code is partially re-installed and all the desired symbols from each transmit antenna are detected with a simple code correlator which correlates to the desired spreading code. Note the descrambling process is also included in the code correlator. Defining an error vector of $\mathbf{z} = \mathbf{d}_i - \mathbf{W}^H \mathbf{y}_{i+F:i-F}$ and an error covariance matrix $\mathbf{R}_{zz} = E[\mathbf{z}\mathbf{z}^H]$, the MIMO LMMSE chip-level equalizer $\mathbf{W}$ is the solution of the following problem:

$$\mathbf{W}^{opt} = \arg\min_{\mathbf{W}} \mathrm{Trace}(\mathbf{R}_{zz}) = \arg\min_{\mathbf{W}} E\|\mathbf{d}_i - \mathbf{W}^H \mathbf{y}_{i+F:i-F}\|^2, \tag{4}$$

whose optimal solution is given by:

$$\mathbf{W}^{opt} = \sigma_d^2 \mathbf{R}^{-1} \mathbf{H}_{i:i}. \tag{5}$$

Where $\mathbf{R} = E[\mathbf{y}_{i+F:i-F}\mathbf{y}_{i+F:i-F}^H]$ is the correlation matrix of the received signal, σ_d^2 is the transmitted chip power. Meanwhile, although **H** is fixed for a given channel realization and is not a function of symbol index *i*, here we use the notation $\mathbf{H}_{i+F:i+1}$, $\mathbf{H}_{i:i}$ and $\mathbf{H}_{i-1:i-F-L}$ to represent the sub-matrices of **H** that are associated with $\mathbf{d}_{i+F:i+1}$, $\mathbf{d}_i$ and $\mathbf{d}_{i-1:i-F-L}$ in the expansion of the matrix-vector product:

$$\begin{aligned}\mathbf{H}\mathbf{d}_{i+F:i-F} &\triangleq \begin{bmatrix} \mathbf{H}_{i+F:i+1} & \mathbf{H}_{i:i} & \mathbf{H}_{i-1:i-F-L} \end{bmatrix} \begin{bmatrix} \mathbf{d}_{i+F:i+1} \\ \mathbf{d}_i \\ \mathbf{d}_{i-1:i-F-L} \end{bmatrix} \\ &= \mathbf{H}_{i+F:i+1}\mathbf{d}_{i+F:i+1} + \mathbf{H}_{i:i}\mathbf{d}_i + \mathbf{H}_{i-1:i-F-L}\mathbf{d}_{i-1:i-F-L}. \end{aligned} \tag{6}$$

Constrained Optimization Based Non-linear LMMSESIC

[0017] Both MIMO LMMSE and MIMO LMMSE-FFT belong to the category of linear equalization methods. In this section, we introduce the non-linear decision feedback to the MIMO LMMSE equalizer to improve the overall receiver performance. The resulting LMMSE-SIC (LMMSE- Successive Interference Cancellation) receiver is illustrated for a 2 Tx, N Rx MIMO system in Figure 4. There are two symbol-detection paths in the block diagram where the first/second path aims at detecting symbols transmitted on the first/second Tx antenna. The First path is similar to that of a conventional MIMO LMMSE receiver, where the chip-equalizer—using frequency domain representation for ease of exposition--$w_1(f)$ is designed such that both the inter-chip interference (ICI) and the spatial co-channel interference (CCI) from the other transmit antenna is suppressed in the estimated chip sequence $\hat{d}_1(f)$:, in order to facilitate the de-spreading and symbol detection $\hat{a}_1(f)$ by restoring the orthogonality between the Walsh codes. Note that only the symbols carried on the desired users' Walsh codes are detected. As shown in Figure 4, these detected symbols from the first path are directly fed-back and used in the symbol detection of the second path.

[0018] The second path of the LMMSE-SIC receiver deviates from that of a conventional MIMO LMMSE receiver in that the chip equalizer $w_2(f)$ does not attempt to directly generate the chip sequence from the second transmit antenna. Instead, it attempts to generate a weighted sum of the chip sequences $\hat{d}_s(f)$ from both transmit antennas while suppressing all the ICIs, i.e,

$$\hat{d}_s(f) = d_2(f) + b_{2,1}d_1(f) + n_2(f). \tag{7}$$

this filter provides complete temporal ICI suppression while keeping a "controlled" residue CCI whose strength is denoted by a design parameter $b_{2,1}$. Furthermore, this design parameter is jointly optimized with the filter coefficients to achieve the best performance. The detailed derivation of this joint optimization is provided in the next section. Since the same set of Walsh codes are used in both Tx antennas, it follows that a simple despreading process gives us the symbol

estimate $\hat{a}_s(f)$ (for a particular Walsh code) that also includes a controlled residue CCI:

$$\hat{a}_s(f) = a_2(f) + b_{2,1}a_1(f) + n_2'(f), \tag{8}$$

where $\sigma^2_{n_2'} = 1/G\sigma^2_{n_2}$ and G is the spreading length. It becomes obvious that assuming correct symbol estimations $\hat{a}_1(f)$ from the first path, the symbol estimate for the second Tx antenna is given by:

$$\hat{a}_2(f) = \hat{a}_s(f) - b_{2,1}\hat{a}_1(f). \tag{9}$$

Soft channel-coded bits can be easily extracted from the symbol estimates $\hat{a}_1(f), \hat{a}_2(f)$ before being sent to the decoder to complete the receiver processing. As shown in a later section, significant performance gain can be achieved with LMMSE-SIC, compared to a conventional LMMSE receiver. Furthermore, the LMMSE-SIC algorithm achieve the performance gains without having to make the impractical assumption of knowing all the Walsh codes. This key advantage makes LMMSE-SIC especially attractive for a fixed voice-data system such as 1X EV-DV or HSDPA and the like, where the desired user usually accounts for only 10-50 percent of the overall transmit power in the cell. We illustrate such a mixed voice-data 1X EV-DV system in Figure 5, where the data user of interest and another data user each assumes about 20 percent of the transmit power, while the rest of the transmit power is assigned to voice users in the cell besides some housekeeping overheads such as pilot, synchronization, etc. Also note that in this example system, data users have a spreading length of 32 while the voice users have a spreading length of 64.

Joint Equalizer/Feedback Weights Optimization

[0019] We mentioned in the 2 Tx MIMO example that the filter coefficients and the feedback weight b_{12} requires joint optimization. In this section, we formulate this joint optimization problem for a general M Tx MIMO

system and show that the problem can be converted into a series of M MVDR problems that are easily solved with the Lagrange Multiplier algorithm.

[0020] For ease of exposition, we return to the matrix-vector notation used in equations (3 -6). The joint equalizer/feedback weights optimization for a M Tx, N Rx MIMO system can be formulated as the following LMMSE problem with a lower-triangular structural constraint on the feedback weights:

$$\mathbf{W}^{opt}, \mathbf{B}^{opt} = \arg\min_{\mathbf{W},\mathbf{B}} \mathrm{Trace}(\mathbf{R}_{zz}) = \arg\min_{\mathbf{W},\mathbf{B}} E\|\mathbf{B}^H \mathbf{d}_i - \mathbf{W}^H \mathbf{y}_{i+F:i-F}\|^2, \quad \text{s.t.} \quad \mathbf{B} = \begin{bmatrix} 1 & & 0 \\ \vdots & \ddots & \\ b_{M,1} & \dots & 1 \end{bmatrix}. \tag{10}$$

[0021] Note that in this case the error vector is defined as $\mathbf{z} = \mathbf{B}^H \mathbf{d}_i - \mathbf{W}^H \mathbf{y}_{i+F:i-F}$. A direct application of Lagrange multipliers to (10) proves to be difficult. To exploit the structure of the problem, we re-organize the structural constraint in (10) into two constraints that are more compact[17]:

$$\begin{aligned} b_{m,m} &= 1, \; m = 1, \cdots, M \text{ and} \\ \mathcal{Z}\mathbf{b} &= 0 \end{aligned} \tag{11}$$

where $b_{m,m} \triangleq (\mathbf{B})_{m,m}$ and

in which $\mathbf{b}_m$ is the m_{th} column of $\mathbf{B}$. Furthermore, Z is defined as

$$\mathcal{Z} \triangleq \begin{bmatrix} \mathcal{Z}_1 & & \\ & \ddots & \\ & & \mathcal{Z}_M \end{bmatrix} \tag{13}$$

where each sub-matrix Z_m is an $M \times M$ diagonal matrix denoting the index of the null elements in $\mathbf{b}_i$, that is, $(Z_m)_{k,k} = 1$ if the k_{th} element in $\mathbf{b}_m$ is constrained to be zero, and $(Z_m)_{k,k} = 0$ if it is not. Similarly, we define another diagonal matrix $N = \mathrm{diag}\{N_1, \ldots, N_M\}$ that denotes the index to the non-zero elements in $\mathbf{b}$. It is easy to see that $Z_m + N_m = \mathbf{I}$ by definition, where $\mathbf{I}$ is the identity matrix.

With these definitions, we set up the MMSE problem to provide optimal $\mathbf{W}$ and $\mathbf{B}$ with a generalized feedback structure:

$$\begin{aligned} \mathbf{W}^{opt}, \mathbf{B}^{opt} &= \arg\min_{\mathbf{W},\mathbf{B}} E\|\mathbf{B}^H \mathbf{d}_i - \mathbf{W}^H \mathbf{y}_{i+F:i-F}\|^2, \quad \text{s.t.} \\ b_{m,m} &= 1, \ m = 1, \cdots, M \ \text{and} \\ \mathcal{Z}\mathrm{b} &= 0. \end{aligned} \tag{14}$$

[0022] To proceed, note that the cost function J in (14) can be re-written as:

$$\begin{aligned} J &= E\|\mathbf{W}^H \mathbf{y}_{i+F:i-F} - \mathbf{B}^H \mathbf{d}_i\|^2 \\ &= \sum_{m=1}^{M} \|\mathbf{w}_m^H \mathbf{H}_{i:i} - \mathbf{b}_m^H\|^2 + \sum_{m=1}^{M} \mathbf{w}_m^H \mathbf{V} \mathbf{w}_m \end{aligned} \tag{15}$$

where $\mathbf{V} \triangleq \sigma_d^2 \bar{\mathbf{H}}_{i:i} \bar{\mathbf{H}}_{i:i}^H + \sigma^2 \mathbf{I}$, $\mathbf{w}_m$ is the mth columns of $\mathbf{W}$. Note that $\bar{\mathbf{H}}_{i:i}$ is a submatrix of $\mathbf{H}$ that is obtained by excluding $\mathbf{H}_{i:i}$ in the matrix $\mathbf{H}$. Meanwhile, in the derivation we have also substituted $\mathbf{y}_{i+F:i-F}$ from (3). We now use the identity $\mathbf{I} = \mathcal{Z}_m + \mathcal{N}_m$ to break up $\mathbf{H}_{i:i}$ and $\mathbf{b}^H_m$ as:

$$\begin{aligned} \mathbf{H}_{i:i} &= \mathbf{H}_{i:i}\mathcal{N}_m + \mathbf{H}_{i:i}\mathcal{Z}_m, \\ \mathbf{b}_m^H &= \mathbf{b}_m^H \mathcal{N}_m + \mathbf{b}_m^H \mathcal{Z}_m. \end{aligned} \tag{16}$$

[0023] Substituting (16) into (15) and after some algebra, we get

$$J = \sum_{m=1}^{M} \|\mathbf{w}_m^H \mathbf{H}_{i:i} \mathcal{N}_m - \mathbf{b}_m^H \mathcal{N}_m\|^2 + \sum_{m=1}^{M} \mathbf{w}_m^H \mathbf{V}_m \mathbf{w}_m \tag{17}$$

where we have defined $\mathbf{V}_m \triangleq \mathbf{H}_{i:i} \mathcal{Z}_m \mathbf{H}_{i:i}^H + \mathbf{V}$. Since $\mathbf{b}_m^H \mathcal{N}_m$ contains the controllable non-zero elements in $\mathbf{b}_m^H$, we minimize the first term in J by setting

$$\mathbf{b}_m^H \mathcal{N}_m = \mathbf{w}_m^H \mathbf{H}_{i:i} \mathcal{N}_m. \tag{18}$$

Meanwhile, we observe that the second sum in J is optimized by individually minimizing

$$\mathbf{w}_m^{opt} = \arg\min_{\mathbf{w}_m} \mathbf{w}_m^H \mathbf{V}_m \mathbf{w}_m; \quad \text{s.t.} \quad \mathbf{w}_m^H \mathbf{h}_m = 1; \quad \text{for } m = 1, \ldots, M \tag{19}$$

[0024] Note that $\mathbf{h}_m$ is the m_{th} column of the matrix $\mathbf{H}_{i:i}$, and the constraint $\mathbf{w}_m^H \mathbf{h}_m = 1$ follows from (18) and the original constraint $b_{m,m} = 1$. We have now decoupled the complex optimization problem (10) into a series of simple MVDR optimization problems (19) to get the filter coefficients $\mathbf{w}_m$. The non-zero elements in the feedback weight vectors $\mathbf{b}_m$ are then obtained from (18). A straightforward application of the Lagrange Multiplier method to (19) leads to the final solution to the joint optimization problem:

$$\begin{aligned} \mathbf{w}_m^{opt} &= \frac{\mathbf{V}_m^{-1}\mathbf{h}_m}{\mathbf{h}_m^H \mathbf{V}_m^{-1}\mathbf{h}_m} \\ (\mathbf{b}_m^{opt})^H \mathcal{N}_m &= (\mathbf{w}_m^H)^{opt} \mathbf{H}_{i:i} \mathcal{N}_m, \end{aligned} \tag{20}$$

and the elements of resulting error correlation matrix is given by

$$(\mathbf{R}_{zz})_{m,n} = (\mathbf{w}_m^{opt})^H (\mathbf{V} + \mathbf{H}_{i:i} \mathcal{Z}_m \mathcal{Z}_n \mathbf{H}_{i:i}^H) \mathbf{w}_n^{opt}. \tag{21}$$

Soft Decision Feedback

[0025] In the 2 Tx, 2Rx example illustrated in section 4, the estimated symbol from the first path $\hat{a}_1(f)$ is used in (9) to obtain the symbol estimates for the second path. If these symbol estimates $\hat{a}_1(f)$ from first path are generated by hard-decisions, then the performance degradation for the second path can be severe due to error propagations.

[0026] The error propagation problem can be alleviated by using the soft decision feedback. There are many types of soft estimators in literature, we found that the so-called conditional mean based estimator to be both analytically pleasing and easy to implement. This conditional mean estimator (also known as MMSE estimator) is also used in similar decision feedback algorithms such as [16]. Consider the general signal model for an arbitrary

symbol at the output of the de-spreader:

$$r_{k,m}(j) = ca_{k,m}(j) + n, \tag{22}$$

where c is a multiplicative factor and k,m, j are Walsh code, transmit antenna and symbol indices, respectively. Also note that we assume n to Gaussian: $n \sim \mathcal{N}(0, \sigma_n^2)$ and both c and σ_n^2 are functions of the spreading gain G, transmit power α_k, noise level σ^2, channel **H** and the filter **W***opt*. Once c and σ_n^2 are obtained, the conditional mean estimate of $a_{k,m(j)}$ is given by

$$\hat{a}_{k,m}(j) = E[a_{k,m}(j) | r_{k,m}(j)], \tag{23}$$

which has a closed-form solution if we invoke the Gaussian assumption on n. For example, when the symbols of interest are QPSK modulated with unit energy, the solution reduces to a simple hyperbolic tangent function:

$$\hat{a}_{k,m}(j) = \frac{1}{\sqrt{2}} tanh \left(\frac{Re(\sqrt{2}c^* r_{k,m}(j))}{\sigma_n^2} \right) + j \frac{1}{\sqrt{2}} tanh \left(\frac{Im(\sqrt{2}c^* r_{k,m}(j))}{\sigma_n^2} \right). \tag{24}$$

[0027] The reason that the conditional mean estimator works well can be intuitively explained by observing Figure 6, where the real part of the symbol estimate $Re(\hat{a}_{k,m}(j))$ is plotted as 10 a function of the real part of the initial noisy soft estimate $Re(r_{k,m}(j))$. The hyperbolic function essentially acts as a clipper, which provides near-hard decision output when the initial estimate is reliable, i.e, when $|Re(r_{k,m}(j))|$ is large. On the contrary, if the initial estimate is less reliable, the output is reduced to alleviate the possible error propagation impact.

Optimization of Detection Order and Further Iterations

[0028] In the LMMSE-SIC algorithm discussed above, we have .xed the detection order to be (1, 2, ...,M) rather arbitrarily. However, the performance

of the successive detection can be improved by optimizing the detection order [4] [5] [18]. In [4], the detection order for a similar V-BLAST problem is chosen so that the worst SNR among M data streams is maximized. In particular, let denote an arbitrary ordering and let Ω be the set of all possible orderings. The cardinality of the set is | Ω | = M! and the following problem is solved to obtain the optimal ordering β

$$\beta = \arg\max_{\omega \in \Omega} \quad \min_{m=1}^{M} \gamma_m(\omega), \tag{25}$$

where $\gamma_m(\omega)$ denotes the effective SNR associated with the detection of the m_{th} transmitted data sequence for a given detection order ω.. The effective SNR can be equivalently defined either at chip-level or at symbol level. The chip-level SNR for the m_{th} data sequence is defined as a function of the error covariance matrix $\mathbf{R}_{zz}$:

$$\gamma_m(\omega) = \frac{\sigma_d^2}{(\mathbf{R}_{zz}(\omega))_{m,m}}, \tag{26}$$

which reduces to the following simple form considering the results in (20) and (21):

$$\gamma_m(\omega) = \sigma_d^2 \mathbf{h}_m^H \mathbf{V}_m^{-1}(\omega)\mathbf{h}_m. \tag{27}$$

[0029] Once the effective SNR is defined, the problem in (25) can be solved by a “localized optimization” procedure similar to the one first proposed in [4]. In a localized optimization procedure, at each stage of the detection process, the sequence that has the best SNR is chosen for detection. The detail of this local optimization is not the focus of our discussion here.

[0030] The performance of the LMMSE-SIC algorithm can be improved if we allow more complexity and introduce further iterations after all the data sequences are detected using the successive algorithm described above. Without loss of generality, we assume that in the .rst iteration the detection order is $\boldsymbol{\omega}$ = (1, 2, · · ·,M) and the filter matrix **W** and feedback matrix **B** is obtained by solving the optimization problem (10). In the second iteration,

however, the design of the matrix pencil **W,B** should be changed to reflect the fact that at the beginning of the second iteration, we have already obtained an initial estimate of all the transmitted data sequences. To this end, we change the constraint in (10) and rewrite the optimization problem as:

$$\begin{aligned} \mathbf{W}^{opt}, \mathbf{B}^{opt} &= \arg\min_{\mathbf{W},\mathbf{B}} E\|\mathbf{B}^H \mathbf{d}_i - \mathbf{W}^H \mathbf{y}_{i+F:i-F}\|^2, \\ \text{s.t.} \qquad \mathbf{B} &= \begin{bmatrix} 1 & \ldots & b_{1,M} \\ \vdots & \ddots & \vdots \\ b_{M,1} & \ldots & 1 \end{bmatrix}. \end{aligned} \tag{28}$$

[0031] It is observed that in (28), none of the elements in the feedback matrix $\mathbf{B}$ is constrained to be zero. Therefore, it allows maximum amount of decision feedback in the spatial dimension and thus improve the effective SNR for each data sequence. Note that since (28) can also be transformed into the form shown in (14) with the compact constraint representation, the solution in (20-21) applies to (28) as well. However, it is easy to see that in this case,

$$\mathcal{Z}_1 = \cdots = \mathcal{Z}_M = \mathbf{0}, \tag{29}$$

and as a result

$$\mathbf{V}_1 = \cdots = \mathbf{V}_M = \mathbf{V}. \tag{30}$$

[0032] Consequently, the solution to (28) assumes a simpler form:

$$\begin{aligned} \mathbf{w}_m^{opt} &= \frac{\mathbf{V}^{-1}\mathbf{h}_m}{\mathbf{h}_m^H \mathbf{V}^{-1} \mathbf{h}_m} \\ (\mathbf{b}_m^{opt})^H &= (\mathbf{w}_m^H)^{opt} \mathbf{H}_{i:i}. \end{aligned} \tag{31}$$

[0033] We further denote γ_m^{II} as the effective SNR for the *mth* data sequence in the second iteration. It is easy to show that:

$$\gamma_m^{II} = \sigma_d^2 \mathbf{h}_m^H \mathbf{V}^{-1} \mathbf{h}_m. \tag{32}$$

[0034] Note that unlike $\mathbf{V}_m$, $\mathbf{V}$ is independent of the detection order. Consequently, both $\mathbf{w}_m$, $\mathbf{b}_m$ and γ_m^{II} are independent of the detection order, meaning that all these quantities can be computed at the same time, unlike the local optimization process required in the first iteration. The detection order in the second iteration is thus obtained by simply sorting all the γ_m^{II} in the descending order. Furthermore, denoting γ_m^{I} as the SNR for m_{th} data sequence in the first iteration, one can easily show that

$$\gamma_m^{I} \leq \gamma_m^{II}, \tag{33}$$

for any 1 $1 \leq m \leq M$. Note that this inequality holds regardless of the detection order used in the first and second iterations. This inequality also signifies the benefit of having a second iteration in the LMMSE-SIC algorithm, provided that the added complexity is within the confine of a practical implementation.

FFT-based Complexity Reduction for a 2 Tx LMMSE-SIC Algorithm

[0035] In [19], an LMMSE-FFT algorithm originally proposed for a SISO/SIMO system in [14] is extended to the MIMO system. It was shown in [19] that the complexity of the an MIMO LMMSE algorithm can be greatly reduced with the FFT-based approach that exploits the block Toepltiz structure of the received signal correlation matrix $\mathbf{R}$, and approximate $\mathbf{R}^{-1}$ with $\mathbf{S}^{-1}$, where $\mathbf{S}$ is the associated block circulant matrix. It will be desirable if a similar lowcomplexity approach exists for the MIMO LMMSE-SIC algorithm as well. Unfortunately, since the matrices to be inverted in the LMMSE-SIC algorithm, namely $\mathbf{V}_m, m = 1, \cdots, M$, are not block Toeplitz, the direct extension of the FFT-based approach to the LMMSE-SIC algorithm is not possible for a general M Tx, N Rx MIMO system. For the special case where the number of transmit antenna $M = 2$, the matrices $\mathbf{V}_1$ and $\mathbf{V}_2$ can be written as a rank-one or rank-two update of a block Toeplitz matrix, which allows a low-complexity inversion that is similar to the MIMO LMMSE-FFT approach.

We focus on the .rst iteration of LMMSE-SIC and assume that the detection order is $\omega = (1,2)$.

One can show that in this case

$$
\begin{aligned}
\mathbf{V}_1 &= \mathbf{V} + \sigma_d^2 \mathbf{h}_2 \mathbf{h}_2^H \\
\mathbf{V}_2 &= \mathbf{V},
\end{aligned}
\tag{34}
$$

and they relate to the correlation matrix $\mathbf{R}$ by

$$
\begin{aligned}
\mathbf{V}_1 &= \mathbf{R} - \sigma_d^2 \mathbf{h}_1 \mathbf{h}_1^H \\
\mathbf{V}_2 &= \mathbf{R} - \sigma_d^2 \mathbf{H}_{i:i} \mathbf{H}_{i:i}^H = \mathbf{R} - \sigma_d^2 \mathbf{h}_1 \mathbf{h}_1^H - \sigma_d^2 \mathbf{h}_2 \mathbf{h}_2^H.
\end{aligned}
\tag{35}
$$

[0036] Substituting (35) into (20) and using the identity that

$$
\frac{\mathbf{A}^{-1}\mathbf{c}}{\mathbf{c}^H \mathbf{A}^{-1}\mathbf{c}} = \frac{(\mathbf{A} - \sigma_d^2 \mathbf{c}\mathbf{c}^H)^{-1}\mathbf{c}}{\mathbf{c}^h (\mathbf{A} - \sigma_d^2 \mathbf{c}\mathbf{c}^H)^{-1}\mathbf{c}},
\tag{36}
$$

we can write the optimal filters as a function of correlation matrix $\mathbf{R}$:

$$
\begin{aligned}
\mathbf{w}_1 &= \frac{\mathbf{V}_1^{-1}\mathbf{h}_1}{\mathbf{h}_1^H \mathbf{V}_1^{-1}\mathbf{h}_1} = \frac{\mathbf{R}^{-1}\mathbf{h}_1}{\mathbf{h}_1^H \mathbf{R}^{-1}\mathbf{h}_1}, \\
\mathbf{w}_2 &= \frac{\mathbf{V}_2^{-1}\mathbf{h}_2}{\mathbf{h}_2^H \mathbf{V}_2^{-1}\mathbf{h}_2} = \frac{(\mathbf{R} - \sigma_d^2 \mathbf{h}_1 \mathbf{h}_1^H)^{-1}\mathbf{h}_2}{\mathbf{h}_2^H (\mathbf{R} - \sigma_d^2 \mathbf{h}_1 \mathbf{h}_1^H)^{-1}\mathbf{h}_2}.
\end{aligned}
\tag{37}
$$

[0037] It is clear in (37) that the FFT-based approach can be used directly to get $\mathbf{w}_1$ by approximating $\mathbf{R}$ with its associated circulant matrix $\mathbf{S}$: $\mathbf{R}^{-1} \approx \mathbf{S}^{-1}$.. On the other hand, the application of the FFT approach to $\mathbf{w}_2$ requires a little more work. Observing that $\mathbf{R} - \sigma_d^2 \mathbf{h}_1 \mathbf{h}_1^H$ rank-one update of the block Toeplitz matrix $\mathbf{R}$, we invoke the matrix inversion Lemma [20] and obtain:

$$
(\mathbf{R} - \sigma_d^2 \mathbf{h}_1 \mathbf{h}_1^H)^{-1} = \mathbf{R}^{-1} + \frac{\sigma_d^2 \mathbf{R}^{-1}\mathbf{h}_1 \mathbf{h}_1^H \mathbf{R}^{-1}}{1 - \sigma_d^2 \mathbf{h}_1^H \mathbf{R}^{-1}\mathbf{h}_1} \approx \mathbf{S}^{-1} + \frac{\sigma_d^2 \mathbf{S}^{-1}\mathbf{h}_1 \mathbf{h}_1^H \mathbf{S}^{-1}}{1 - \sigma_d^2 \mathbf{h}_1^H \mathbf{S}^{-1}\mathbf{h}_1}.
\tag{38}
$$

[0038] Note that in (38) we have already used the approximation $\mathbf{R}^{-1} \approx \mathbf{S}^{-1}$. From (37 -38), we have shown that for a 2Tx MIMO system using LMMSE-SIC algorithm, the filters $\mathbf{w}_1$ and $\mathbf{w}_2$ can be obtained using FFT-

based approach to compute the inversion of the associated circulant matrix $\mathbf{S}$, instead of the direct inversion of the correlation $\mathbf{R}$. As a result, the complexity of the MIMO LMMSE-SIC algorithm for a 2Tx system can be close to that of the MIMO LMMSE-FFT algorithm, provided that no further iterations are used in the LMMSE-SIC.

Simulation Results

[0039] The proposed MIMO LMMSE-SIC algorithms is evaluated in a realistic simulation chain, and the simulation parameters are tabulated in Table I.

Parameter Name	Parameter Value
System	CDMA 1X/EVDV
Spreading Length	32
Channel Profile	Vehicular A
Mobile Speed	50 km/h
Filter Length	32
Number of Tx/Rx Antennas	2/2
Modulation Format	QPSK
Information Data Rate	312 kbps
Turbo Code Rate	0.6771
Geometry	6
Number of Walsh Codes Assigned to the user	3
Total number of Active Walsh Codes in the system	25

Table 1: Simulation parameters.

[0040] Note that since the project is ongoing, the 14 results presented here are not complete and more results will be included in a future revision of this document. Nevertheless, these results provide good insights about the effectiveness of the algorithm proposed in this report. The benefit of the non-linear decision feedback is demonstrated in Figure 7, where the MIMO LMMSE-SIC algorithm is compared with the conventional LMMSE algorithm. Note that the impact of detection order optimization is included in these results. There are three LMMSE-SIC curves shown in this figure, representing the hard-decision feedback method, the conditional mean estimator based soft decision feedback method and the ideal feedback case. It is observed that even with the hard decision feedback, the MIMO LMMSE-SIC algorithm outperform the conventional LMMSE by about 1 dB at high SNR. Another 0.5-1dB can be gained if we replace the hard decisions with the soft decisions generated by the conditional mean estimator. It is important to note that these

gains are achieved with the assumption that the mobile receiver has only the knowledge of 3 out of 25 active Walsh codes.

[0041] In Figure 8, the impact of further iterations in the LMMSE-SIC algorithm is investigated using the same example. Note that for both one iteration an two iterations case, the conditional mean estimator based soft decision feedback are used. In this case, adding a second iteration actually degrades the performance at high SNR. This detrimental effect, also known as "ping-pong" effect in the iterative detectors, is mainly attributed to the accumulation of decision errors in the iterative process. Obviously, for the MIMO LMMSE-SIC algorithm, further research is needed to reduce the impairment caused by the accumulative error propagation.

[0042] Embodiments of the invention describe a MIMO receiver design that provides the best performance and complexity trade off. An embodiment of the invention is a potential candidate reference receiver for MIMO standards for 3GPP2.

[0043] Although described in the context of particular embodiments, it will be apparent to those skilled in the art that a number of modifications and various changes to these teachings may occur. Thus, while the invention has been particularly shown and described with respect to one or more preferred embodiments thereof, it will be understood by those skilled in the art that certain modifications or changes, in form and shape, may be made therein without departing from the scope and spirit of the invention as set forth above.

ABSTRACT

[0044] A system according to embodiments of this invention provide a multiple transmit antenna, multiple receive antenna (MIMO) receiver design for the communication downlinks such as those used in CDMA technology. Two algorithms referred to as the MIMO LMMSE-FFT and MIMO LMMSE-SIC (Successive Interference Cancellation) algorithms, are described in detail. In embodiments of the invention, the interference cancellation step is achieved without the impractical assumption of the knowledge of all the active Walsh codes in the systems, unlike many other interference cancellation based algorithms found in the literature.

1/8

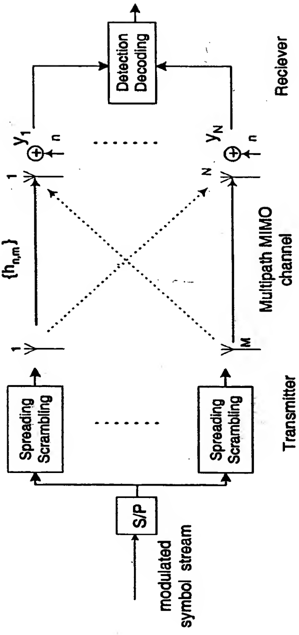


Figure 1: MIMO-CDMA system illustrated.

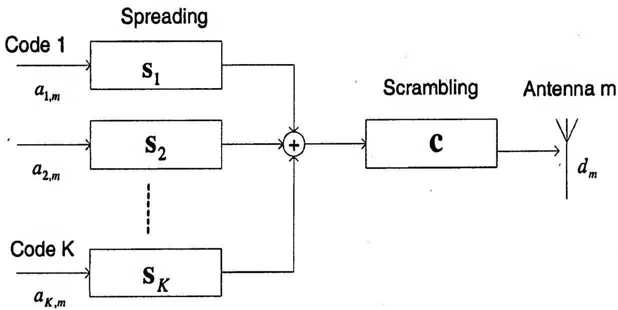


Figure 2: Transmit signaling at antenna m.

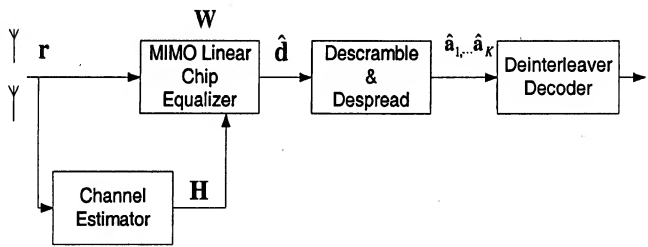
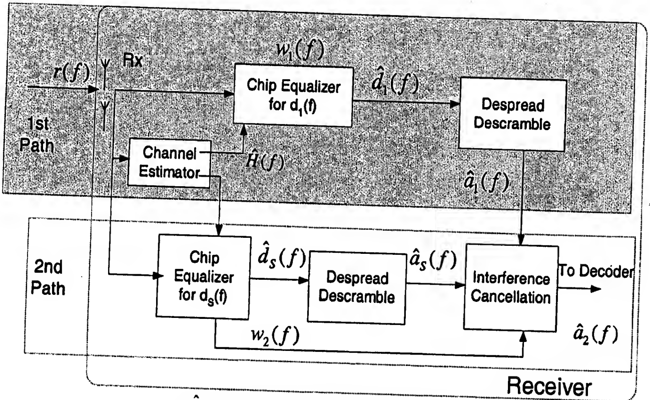


Figure 3: Block diagram of the MIMO-LMMSE algorithm.



- $\hat{d}_S(f)$ includes weighted sum of $d_1(f)$ and $d_2(f)$

Figure 4: Block diagram of an LMMSE-SIC receiver.

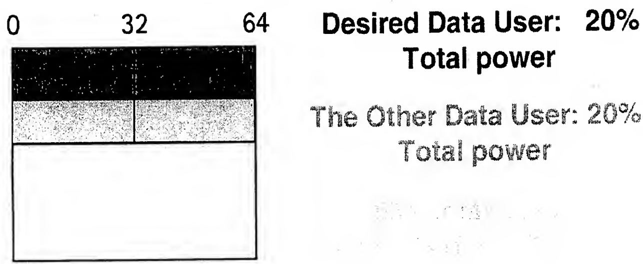


Figure 5: Illustration of a mixed traffic 1X EV-DV system.

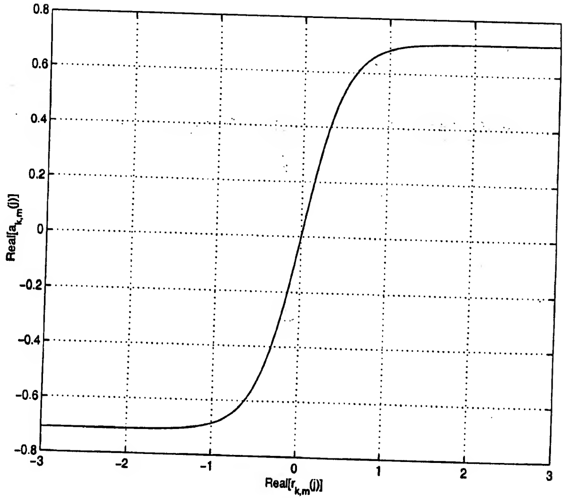


Figure 6: Illustration of the conditional mean estimator.

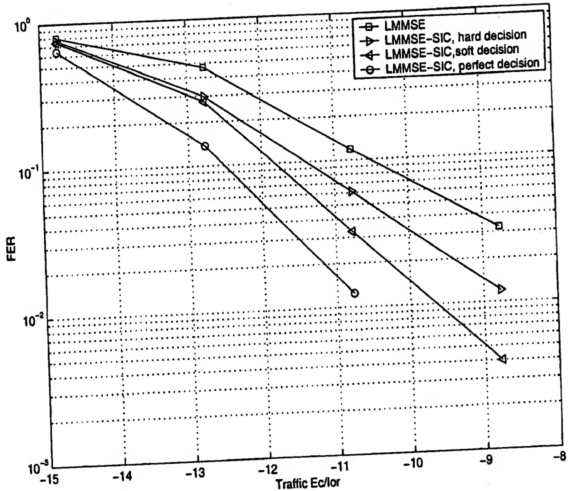


Figure 7: Simulation results for the MIMO LMMSE-SIC algorithm.

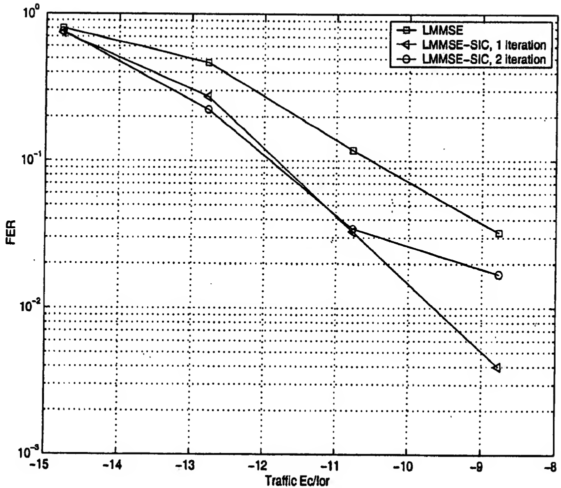


Figure 8: Simulation results for the MIMO LMMSE-SIC algorithm with 2 iterations.